Realization of Current control strategies of Shunt Active Power Filter under Unbalanced Load

A THESIS SUBMITTED IN PARTIAL FULFILLMENT
OF THE REQUIREMENTS FOR THE DEGREE OF

Master of Technology

In

Power Control and Drives

By

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Under the Guidance of

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CERTIFICATE

This is to certify that the thesis entitled**, "REALIZATION OF CURRENT CONTROL STRATEGIES OF SHUNT ACTIVE POWER FILTER UNDER UNBALANCED LOAD"** submitted by Mr. **Narayana Divakar.R.V.L** in partial fulfillment of the requirements for the award of Master of Technology in Electrical Engineering with specialization in **"Power Control and Drives"** during session 2011-13 at the National Institute of Technology, Rourkela is an authentic work carried out by him under my supervision and guidance.

To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other University/ Institute for the award of any degree or diploma.

Date: **Prof. B. Chitti Babu**

Place: Rourkela **Supervisor**

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Narayana Divakar.R.V.L

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ABSTRACT

Filtering is the major criteria to maintain power quality improvement in the three phase three wire system. Now a day Active power Filters (APF) is mostly preferable for this purpose. Conventional PI control methods are disadvantageous in eliminating high harmonics in the line current and 2^{nd} harmonics at the DC link of a shunt APF. This PR control method which nullifies 2^{nd} harmonics and provides better performance and accuracy compared to the Conventional positive-sequence control method and DC link voltage control method individually in terms of power quality improvement and cost effectiveness. The performance of the three control methods are demonstrated through simulation and experimental results.

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ABBREVIATIONS USED

APF	Active Power Filter
PI Controller	Proportional Integral Controller
PR Controller	Proportional Resonant Controller
SSI	Sinusoidal Signal Integrator
THD	Total Harmonic Distortion

VARIABLES USED

d_{dq}	Switching function of the inverter
I_{dc}	Current in the DC link
V_{dc}	DC link voltage
ε_2	Error of the Voltage control loop
ω_e	Natural frequency
ω_c	Cutoff frequency
K_p	Proportional gain
K_i	Integral gain
$G_{\alpha\beta}$	PR Transfer function in terms of Stationary reference frame
$G_{\alpha\beta}^{+1}$	Positive sequence component of PR Transfer function
$G_{\alpha\beta}^{-1}$	Negative sequence component of PR Transfer function
G_{dq}	PI Transfer function in terms of Rotating reference frame
G_{dq1}	Positive sequence component of PI Transfer function
G_{dq2}	Negative sequence component of PI Transfer function

CHAPTER 1

INTRODUCTION

1. INTRODUCTION

Power quality has become a major research topic in power distribution systems due to a significant increase of harmonic pollution caused by proliferation of nonlinear loads such as diode rectifiers, switching power supplied and other types of line connected power converters etc. The shunt APF is recognized as a cost effective solution for harmonic compensation in low and medium power systems. It composed of a PWM voltage source inverter, with a large dc link capacitor, and connected to the line by means of an inductor [1].

Now a day most of the loads are nonlinear due to usage of power electronic devices like semiconductor devices used in rectifiers and inverters, switching power supply and other power electronic converters etc... To overcome above problems shunt APF is recognized as cost effective solution for compensating harmonics in low and medium power applications.

In general, PI controllers are played major role for controlling shunt APF. For three phase Systems synchronous frame PI controllers can be used but it requires computational burden in case of multiple frame transformations. To overcome the above problem Proportional resonant controller is used. The frequency response characteristic of a PR controller is similar to that of a PI controller. PR controllers are used for reference tracking in the stationary reference frame. The basic functionality of a PR controller is to introduce infinite gain at a selected resonant frequency for eliminating steady state error at that frequency.

Basically, an integrator whose DC gain forces the DC steady state error to zero in the same way resonant portion of the PR controller whose AC gain (GI) forces the AC steady error to zero. All the advantages will give importance to PR controller.

Under unbalanced load the fundamental negative sequence component appears in load current. It can result in the fundamental negative sequence component in the APF AC currents and APF AC voltages. The interaction of negative and positive sequence component of switching functions and ac currents of the APF produces a 2^{nd} order harmonic ripple on the DC link of the APF, which will generate the 3rd order harmonic distortion in AC currents of the APF and line currents [2]. In addition, under a nonlinear load ac current of the APF contains high harmonics due to load current harmonics. The interaction of fundamental negative and positive sequence component of switching functions and high harmonics AC currents of APF produces high order even harmonics on the DC link voltage of the APF, which create high order odd harmonics in the

APF AC currents. Such harmonics flow into the line, thereby lead to worsen the performance of the system. Moreover, the high harmonic voltage across capacitors in dc side results in high temperature due to their ESR. This leads to reduce their life-cycle of the capacitors.

Some methods are proposed in the papers [2]-[4] to improve performance of the system. One method is to eliminate the 3rd harmonic of the APF ac current, thus to reduce the distortion of the line current. With this method only 3rd harmonic of APF current can be cancelled, the others high harmonics cannot be reduced. Moreover the even harmonics on the DC link side still exist. In [2] a DC link voltage control method is proposed. Although this control strategy can almost cancel the DC link ripple, the controller is complex.

In order to avoid the disadvantages of the above control method and to eliminate a series harmonics at both DC side and AC side of the APF, PR control method is proposed. This method is implemented with a simple controller.

1.1 LITERATURE REVIEW:

At the end of 19^{th} century the development of alternating current (ac) transmission system was based on sinusoidal voltage at constant frequency generation. Sinusoidal voltage with constant frequency has made easier the design of transformer, machines and transmission lines. If the voltage will be non-sinusoidal then it will create many complications in the design of transformer, machine and transmission system. Nevertheless, some papers were published in the 1920s, showing the conventional concept of reactive and apparent power losses its usefulness in non-sinusoidal cases. Then, two important methods to power definitions under non sinusoidal condition were introduced by Budeanu in 1927 and Fryze in 1932.Fryze defined power in time domain whereas Budeanu did it in frequency domain. Subsequently power electronics was introduced in 1960s; non-linear loads that consume non sinusoidal current have increased significantly. Today everywhere power electronics based equipment are used from domestic purpose to residential purpose. In 1976, Harshima, Inaba and Tsuboi presented, probably for the first time, the term "instantaneous reactive power" for a single phase circuit. That same year, Gyugyi and Strycula used the term "active ac filters" for the first time. A few years later, in 1981, Takahashi, Fujiwara, and Nabae published two papers giving the hint of the appearance of the instantaneous power theory or "p-q theory" [6].

The Power quality is defined as “any incidence established in voltage, current, or frequency deviations that results in failure, damage, upset, or disoperation of end-use equipment.” Today, most of the power quality issues are related to the power electronics equipment which is used in commercial, domestic and industrial application. Therefore, the applications of power electronics equipment for residential purposes like TVs, PCs, Refrigerator etc., business purposes like copiers, printers etc., industrial purposes like PLCs (Programmable logic controller), ASDs (Adjustable speed drive), rectifiers, inverters etc. Today almost all electrical equipment is based on power electronics which causes harmonics, inter-harmonics, notches and neutral currents. Transformers, motors, cables, interrupters, and capacitors (resonance) are some of the equipment which is affected by harmonics. Notches are produced mainly because of the converters, and they basically affect the electronic control devices. Currents in neutral wire are produced in that equipment which uses switched-mode power supplies, such as printers, photocopiers, PCs, and any triplets' generator. Neutral current disturbs the neutral conductor temperature and transformer capability. Inter-harmonics are produced because of cyclo-converters, static frequency converters, arching devices and induction motors.

The existence of harmonics in the power lines results in greater power losses in distribution, and cause problem by interfering in communication systems and, sometime cause operation failures of electronic equipment, which are more and more critical because it consists of microelectronic control systems, which work under very low energy levels. Because of these problems, the power quality issues delivered to the end consumers are of great concern. International standards concerning electrical power quality *(IEEE*-519, *IEC* 61000, *EN* 50160, etc.) impose that electrical equipment should have limitation on the injection of harmonics in 6 the system within a specified limit which has been satisfied by the international standards. Meanwhile, it is very important to solve the problems of harmonics caused by that equipment which is already installed. The major causes of power quality problems are due to the wide spread application of nonlinear loads such as fluorescent lamps, saturable devices, static power electronic converters and arch furnaces. This equipment draws harmonic and reactive power components of current from the ac mains. In three phase system, they can cause unbalance and draw excessive neutral currents. The injected harmonic, reactive power burden, and excessive neutral currents cause low system efficiency and poor power factor, they also cause disturbance to other consumers.

So far to come out of these problems shunt passive filters (consist of tuned LC filters and/or high pass filters) have been used to improve power factor and to reduce harmonics in power systems. But, shunt passive filters was not giving desired performance which leads to the development of "Active Power Filters (APF's)".

Passive filters have been used as a solution to solve harmonic current problems, but passive filters having many disadvantages, namely: they can filter only the frequencies they were previously tuned for; their operation cannot be limited to a certain load; resonances can occur because of the interaction between the passive filters and other loads, with unpredictable results. To come out of these disadvantages, recent efforts are concentrated in the development of active filters. Control strategy (Bhim 1999 and Joao 2001) is the heart of Active Power Filters which are classified into shunt, series, and combination of both. Mainly active power filters can be classified into following configuration: 1) Shunt configuration in which the filter is connected in Parallel with harmonic loads and 2) Series configuration in which the filter is connected in series with the loads. Taking the basic idea of harmonic cancellation, shunt active filter injects current to directly cancel polluting current while, series active filter compensate the voltage distortion caused by non-linear loads. The performance of active filter is dependent on two parts: current control system and harmonic reference generation. The development of compensating signals in terms of voltages or currents is the important part of APF's control strategy which affects its ratings and transient as well as steady state performance. The control strategy which generates compensating signals is based on time domain or frequency-domain. The frequency domain approach takes the use of the Fourier transform and its analysis, which leads to a large amount of calculations, making the control method much more complicated. In the time domain approach, traditional concepts of circuit analysis and algebraic transformations associated with changes of reference frames are used, simplifying the control task. One of the time domain control strategies is the instantaneous reactive power theory based (p-q theory) control strategy which proposed by Akagi et al. (Hirfumi 1983 and Hirfumi1984) and instantaneous active and reactive current component (id-iq) method. And since (Joao 2003) the p-q theory is based on the time domain, this theory is valid both for steady-state and transient operation, as well as for generic voltage and current waveforms, allowing the control of APF in the real-time; another advantage of this theory is the simplicity of its calculations, since only algebraic operations are required. Voltage source converters are used as the active power filter topologies, which have a DC capacitor voltage control as an energy storage device. Converts a DC voltage into an AC voltage by getting appropriate gating signal the power semiconductor switches. Although a single pulse for each half cycle can be applied to synthesize an AC voltage, for most of the application which shows dynamic performance, pulse width modulation (PWM) is the most commonly used today. PWM techniques applied to a voltage source inverter consist of chopping the DC bus voltage to produce an AC voltage of an arbitrary waveform. With PWM techniques, the ac output of the filter can be controlled as a current or voltage source device [6].

1.2 OBJECTIVE OF WORK DONE:

The overall objectives to be achieved in this study are:

1. Modeling and analysis of dc voltage control method and conventional positive sequence control method for APF system.
2. Modeling and analysis of proposed proportional resonant control model for APF.
3. Compare above three control methods with their THD of line current and 2nd harmonic component of DC link Voltage.

1.3 ORGANIZATION OF THE THESIS:

The organization of the thesis as follows

Chapter 2: presents the mathematical modeling of the shunt active power filter as voltage source inverter for eliminating harmonic content from the line current.

Chapter 3: Presents current control strategies of shunt active power filter for eliminating the 2nd harmonic voltage from the DC link voltage. It reduces the 3rd harmonic current in the line side of the APF. Mathematical modeling of the PR controller is also discussed in this chapter.

Chapter 4: Presents the general conclusion and future work followed by references.

CHAPTER 2

ACTIVE POWER FILTER

2. ACTIVE POWER FILTER:

In [9] an inverter operates as active inductor at a certain frequency to absorb the harmonic current. But the exact calculation of network inductance in real-time is difficult and may deteriorate the control performance. The shunt active power filter acts as active conductance to damp out the harmonics in distribution network [10].The study of the Shunt APF can be classified into three categories;

1. **Harmonic detection:** Generally Shunt APF is connected across the load for compensation purpose. But now a day nonlinear loads are increased due to the usage of power electronic devices. These loads will produce higher order harmonics into the line current. The harmonic detection method to calculate the reference currents of the shunt active power filter.

2. **Structure of Shunt APF:** The Voltage source Inverter with six IGBTs is used for the Shunt APF. Switching pulses of six IGBTs can be generated by the PWM / Hysteresis controller.

3. **PWM / Hysteresis control:** This controller can be used to control the compensating currents. There are many techniques to control the compensating currents such as the Hysteresis current control, the Pulse width modulation control.

2.1 OBJECTIVE OF HARMONIC COMPENSATION:

Objectives of harmonic compensation are as follows;

1. Eliminate real power oscillations
2. Power factor improvement
3. Eliminate current harmonics
4. Provide harmonic damping

These objectives can be done by configure the Active Power Filter

2.2 ACTIVE POWER FILTER CONFIGURATION:

AF's can be classified based on converter type, topology, and the number of phases. The converter type can be either CSI or VSI bridge structure. The topology can be shunt, series, or a combination of both. The third classification is based on the number of phases, such as two-wire (single phase) and three- or four-wire three-phase systems [7].

Coming to the unbalanced load issue, the presence of fundamental negative sequence component in the load currents can affect the APF system performance. This negative sequence, if not blocked by the APF current reference generator, will produce a fundamental negative sequence component in the APF currents due to the proportional gain of the current controller. That will produce a 2^{nd} order ripple in the DC-link voltage, which will generate harmonic distortion in the source currents. Therefore, if the unbalanced load compensation is not required, the circulation of the fundamental negative sequence currents in the APF must be completely blocked to avoid distortion of the source currents [2].

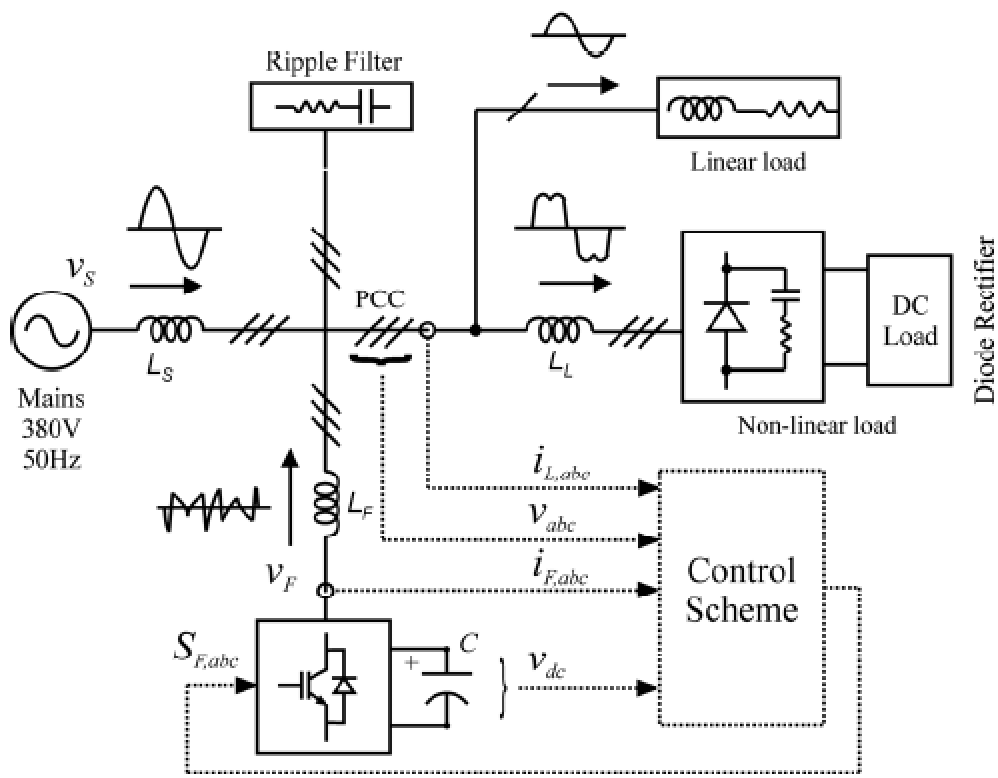

Fig 2.1 Basic current harmonic compensation scheme of an unbalanced load using a shunt APF[2].

2.3 MATHEMATICAL MODEL

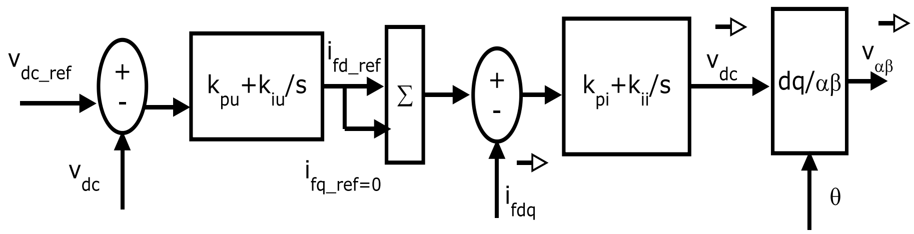

Fig 2.2 The conventional PI control diagram

Continuous-time dynamic model of the Shunt APF system in Fig 2.1 can be represented by equations on the AC and DC side are as follows

$$L_T.\frac{d\bar{i}_F(t)}{dt}=\bar{v}_s(t)-\bar{v}_F(t)-R_T.\bar{i}_F(t) \tag{2.1}$$

$$C.\frac{d\bar{v}_{dc}(t)}{dt}=\bar{i}_{dc}(t) \tag{2.2}$$

Where $L_T = (L_F+L_s)$, $R_T=(R_F+R_S)$, grid voltage, APF poles voltage and APF input current are $\bar{v}_s(t),\bar{v}_F(t),\bar{i}_F(t)$. The inverter pole power $P_{poles}(t)$ can be obtained as

$$P_{poles}(t)=\frac{3}{2}\mathrm{Re}\{S_{poles}(t)\}=\frac{3}{2}\mathrm{Re}\{\bar{v}_F(t).\bar{i}_F^{\,*}(t)\} \tag{2.3}$$

Here subscript “*” represents complex conjugate. Continuous switching vector d(t) can be defined as

$$\bar{d}(t)=\frac{2}{3}(s_a(t)+\alpha.s_b(t)+\alpha^2.s_c(t)) \tag{2.4}$$

Where s_a, s_b, s_c are the duty cycles of each converter leg and $\alpha = e^{j2\pi/3}$. The inverter poles voltage vector is as follows

$$\bar{v}_F(t)=\frac{1}{2}\bar{d}(t)v_{dc}(t) \tag{2.5}$$

Substituting (2.5) in (2.3) yields the active power $P_{poles}(t)$ as

$$P_{poles}(t) = \frac{3}{2}\mathrm{Re}\{\frac{1}{2}\bar{d}(t).v_{dc}(t).\bar{i}_F^{\,*}(t)\} \tag{2.6}$$

After neglecting switching and conduction losses, the active power at the inverter output to the active power at the inverter poles:

$$P_{out} = v_{dc}(t).i_{dc}(t) = P_{poles}(t) = \frac{3}{4}.v_{dc}(t)\,\mathrm{Re}\{\bar{d}(t).\bar{i}_F^{\,*}(t)\} \tag{2.7}$$

So, the i_{dc} current is obtained from (2.7) as

$$i_{dc}(t) = \frac{3}{4}\mathrm{Re}\{\bar{d}(t).\bar{i}_F^{\,*}(t)\} \tag{2.8}$$

Referring (2.8) in stationary (α, β) frame yields

$$i_{dc}(t) = \frac{3}{4}\mathrm{Re}\{\bar{d}_{\alpha\beta}(t).\bar{i}_{F\alpha\beta}^{\,*}(t)\} \tag{2.9}$$

Where

$$\bar{d}_{\alpha\beta} = \bar{d}^{1p}{}_{\alpha\beta} + \bar{d}^{1n}{}_{\alpha\beta} + \sum_{h=6k+1}^{\infty}\bar{d}^{hp}{}_{\alpha\beta} + \sum_{h=6k-1}^{\infty}\bar{d}^{hn}{}_{\alpha\beta} \qquad \text{k=1,2,3………}$$

(2.10)

$$\bar{i}_{F\alpha\beta} = \bar{i}^{1p}{}_{F\alpha\beta} + \bar{i}^{1n}{}_{F\alpha\beta} + \sum_{h=6k+1}^{\infty}\bar{i}^{hp}{}_{F\alpha\beta} + \sum_{h=6k+1}^{\infty}\bar{i}^{hn}{}_{F\alpha\beta} \quad \text{k=1,2,3………} \tag{2.11}$$

Let us consider generic variable $\bar{x}$, space vector of a positive and negative h-order harmonic frequency can be expressed in stationary frame as $\bar{x}^{hp}{}_{\alpha\beta} = e^{jh\omega_e t}.\bar{x}^{hp}{}_{dq}$ (where $\omega_e = 2\pi f_{grid}$ rad/s) and $\bar{x}^{hn}{}_{\alpha\beta} = e^{jh\omega_e t}.\bar{x}^{hn}{}_{dq}$. The vector space vector $\bar{x}^{hp}{}_{dq}$ is the positive h-order harmonic frequency represented in a synchronous frame rotating at positive $h\omega_e t$ and the vector space vector $\bar{x}^{hn}{}_{dq}$ is the negative h-order harmonic frequency represented in a synchronous frame rotating at positive $h\omega_e t$. For our simplicity we are considering that the APF compensates only the first harmonic pair (5^{th}, 7^{th}).

Therefore, the expressions (2.10) and (2.11) can be rewritten as

$$\begin{aligned}\bar{d}_{\alpha\beta} &= \bar{d}^{1p}{}_{\alpha\beta} + \bar{d}^{1n}{}_{\alpha\beta} + \bar{d}^{5n}{}_{\alpha\beta} + \bar{d}^{7p}{}_{\alpha\beta} \\ &= e^{j\omega_e t}.\bar{d}^{1p}{}_{dq} + e^{-j\omega_e t}.\bar{d}^{1n}{}_{dq} + e^{-j\omega_e t}.\bar{d}^{5n}{}_{dq} + e^{j\omega_e t}.\bar{d}^{7p}{}_{dq}\end{aligned} \tag{2.12}$$

In the same way we can represent $i_{F\alpha\beta}$.

$$\begin{aligned}\bar{i}_{F\alpha\beta} &= \bar{i}^{1p}{}_{F\alpha\beta} + \bar{i}^{1n}{}_{F\alpha\beta} + \bar{i}^{5n}{}_{F\alpha\beta} + \bar{i}^{7p}{}_{F\alpha\beta} \\ &= e^{j\omega_e t}.\bar{i}^{1p}{}_{Fdq} + e^{-j\omega_e t}.\bar{i}^{1n}{}_{Fdq} + e^{-j\omega_e t}.\bar{i}^{5n}{}_{Fdq} + e^{j\omega_e t}.\bar{i}^{7p}{}_{Fdq}\end{aligned} \tag{2.13}$$

Replacing (2.12) and (2.13) in (2.9) yields

$$\begin{aligned}i_{dc} &= \frac{3}{4}\mathrm{Re}\{(e^{j\omega_e t}.\bar{d}^{1p}{}_{dq} + e^{-j\omega_e t}.\bar{d}^{1n}{}_{dq} + e^{-j\omega_e t}.\bar{d}^{5n}{}_{dq} + e^{j\omega_e t}.\bar{d}^{7p}{}_{dq}). \\ &(e^{j\omega_e t}.\bar{i}^{1p}{}_{Fdq} + e^{-j\omega_e t}.\bar{i}^{1n}{}_{Fdq} + e^{-j\omega_e t}.\bar{i}^{5n}{}_{Fdq} + e^{j\omega_e t}.\bar{i}^{7p}{}_{Fdq})^{*}\}\end{aligned} \tag{2.14}$$

After expanding the terms and rearranging the real part of (2.14) the DC side current becomes

$$\begin{aligned}i_{dc} &= \frac{3}{4}(I_{dc} + I_{s2}\sin 2\omega_e t + I_{c2}\cos 2\omega_e t + I_{s4}\sin 4\omega_e t + \\ &I_{c4}\cos 4\omega_e t + I_{s6}\sin 6\omega_e t + I_{c6}\cos 6\omega_e t + I_{s8}\sin 8\omega_e t + \\ &I_{c8}\cos 8\omega_{et} t + I_{s12}\sin 12\omega_e t)\end{aligned} \tag{2.15}$$

Where

$$I_{dc} = d^{1p}{}_d.i^{1p}{}_d + d^{1p}{}_q.i^{1p}{}_q + d^{1n}{}_d.i^{1n}{}_d + d^{1n}{}_q.i^{1n}{}_q \tag{2.16}$$

$$I_{s2} = d^{1p}{}_d.i^{1n}{}_q - d^{1p}{}_q.i^{1n}{}_d - d^{1n}{}_d.i^{1p}{}_d + d^{1n}{}_q.i^{1p}{}_d \tag{2.17}$$

$$I_{c2} = d^{1p}{}_d.i^{1n}{}_d + d^{1p}{}_q.i^{1n}{}_q + d^{1n}{}_d.i^{1p}{}_d + d^{1n}{}_q.i^{1p}{}_q \tag{2.18}$$

$$I_{s4} = d^{1n}{}_d.i^{5n}{}_q + d^{5n}{}_d.i^{1n}{}_q \tag{2.19}$$

$$I_{c4} = d^{1n}{}_q.i^{5n}{}_q + d^{5n}{}_d.i^{1n}{}_d \tag{2.20}$$

$$I_{s6} = d^{1p}{}_d.i^{5n}{}_q - d^{1p}{}_d.i^{7n}{}_q - d^{5n}{}_d.i^{1p}{}_q + d^{7p}{}_q.i^{1p}{}_q \tag{2.21}$$

$$I_{c6} = d^{1p}{}_q . i^{5n}{}_q + d^{1p}{}_q . i^{7p}{}_q + d^{5n}{}_d . i^{1p}{}_d + d^{7p}{}_d . i^{1p}{}_q \qquad (2.22)$$

$$I_{s8} = d^{7p}{}_d . i^{1n}{}_q - d^{1n}{}_d . i^{7p}{}_q \qquad (2.23)$$

$$I_{c8} = d^{1n}{}_q . i^{7p}{}_q + d^{7p}{}_d . i^{1n}{}_d \qquad (2.24)$$

$$I_{s12} = d^{7p}{}_d . i^{5n}{}_q - d^{5n}{}_d . i^{7p}{}_q \qquad (2.25)$$

These equations are taken from [2]

Replace i_{Fdq} with i_{dq} in equation (2.16) and write it in normal $v.i^*$ form, we will get equation as follows

Under unbalanced load, the average current through DC-link of the APF can be solved as

$$I_{dc} = \frac{3}{2}[\mathrm{Re}\{\vec{d}_{dq}^{\;+1} . \vec{i}_{dq}^{\;+1^*} + \vec{d}_{dq}^{\;-1} . \vec{i}_{dq}^{\;-1^*}\} + \mathrm{Re}\{e^{j\omega t} . (\vec{d}_{dq}^{\;+1} . \vec{i}_{dq}^{\;-1^*})\} + \mathrm{Re}\{e^{-j\omega t} . (\vec{d}_{dq}^{\;-1} . \vec{i}_{dq}^{\;+1^*})\} \qquad (2.26)$$

According to the mathematical model for the APF and (2.26), DC-link voltage is obtained as

$$V_{dc} = \frac{3}{4C}[I_{dc} + \frac{I_2}{2\omega_e}\sin(2\omega_e t - \alpha_2)] \qquad (2.27)$$

$$u_{dc} = U_{dc} + A_{u_{dc}} \sin(2\omega_e t + \alpha_{u_{dc}}) \qquad (2.28)$$

Where d_{dq}^{+1}, $d_{dq}^{-1}$are vector of switching function positive and negative sequence in the rotating synchronous frame, respectively i_{dq}^{+1*} , i_{dq}^{-1*} are conjunction vector of positive and negative sequence of APF currents in the rotating synchronous frame, respectively, i_{dc} is current through DC link, u_{dc} is DC link voltage of APF. U_{dc} is DC component of u_{dc} , A_u and u_{dc} are magnitude and phase of the 2nd order ripple component of u_{dc} , respectively. It can be seen that the 2nd harmonic ripple appears on the DC link terminal due to the interaction of negative and positive sequence component of input currents with switch functions. Because DC-link voltage contains the 2nd harmonic in (2.27), the output of the PI controller of DC link voltage, i.e., the reference for APF current in the current loop, can be given as (DC item is neglected).

$$\vec{i}_d^{\,*} = A_2\varepsilon_2 \sin(2\omega_e t + \beta_i) = \frac{A_i}{2}(e^{j(2\omega_e t+\beta_i)} - e^{-j(2\omega_e t+\beta_i)}) \tag{2.29}$$

$$\vec{v}_{dq} = A_v \sin(2\omega_e t + \beta_v) = \frac{A_v}{2}(e^{j(2\omega_e t+\beta_v)} - e^{-j(2\omega_e t+\beta_v)}) \tag{2.30}$$

Transforming (2.30) into the form in αβ stationary frame, equation (2.30) is written as

$$\vec{v}_{\alpha\beta} = \vec{v}_{dq} e^{j\omega_e t} = \frac{A_v}{2}(e^{j(3\omega_e t+\beta_v)} - e^{-j(\omega_e t+\beta_v)}) \tag{2.31}$$

In (2.31) the third order harmonic appears in the APF current. Such harmonic results in a fourth order harmonic on the DC link voltage and DC link current of the APF, which in turn causes higher order harmonics in APF ac currents. These harmonics flow into and distort line currents and thereby deteriorate the performance of the system. In addition under nonlinear load, load current contains not only the fundamental negative sequence component, but also high harmonics $6k\pm1$, k= 0, 1, 2... . Interaction of these high harmonics and fundamental positive, negative sequence component leads to a series of high harmonics $2k$, k= 0,1, 2… at DC link, e.g. 2nd, 4th, 6th, 8th etc [5].

CHAPTER 3

CONTROL STRATEGY

3. CONTROL STRATEGY:

Control strategy is to eliminate 2nd harmonic at the DC link voltage of the APF under an unbalanced load condition. As from above analysis, the fundamental negative sequence component is the reason for the production of high harmonics in the line current as well as DC side of the APF. The fundamental Negative sequence component of APF has been reduced to cancel the harmonics in the system. In order to realize this objective, we have to maintain quadrature component of fundamental positive sequence reference current, which is the output of PI controller of voltage control loop, and both the direct and quadrature component of fundamental negative sequence reference current must be zero.

$$i_d^{+1*} = 0 \text{ and } i_{dq}^{-1*} = 0 \qquad (3.1)$$

This is the basis for a PR control method of the APF. The new control method block diagram for the APF system is shown in Fig 3.4. It is having two control loops in the overall system. Outer loop is the voltage control loop, which regulates the DC link voltage of the APF to the reference value. The inner loop includes the fundamental current controller and high harmonic controller. In case of conventional positive sequence control method and DC link voltage control method PI controller is used in fundamental current controller and PR controller is used in high harmonic current controller.

In case of PR control method PR controller is used in both the fundamental sequence current controller and high harmonic current controller. This control method will reduce the complexity of the circuit. The PR control method is compared with the conventional control method and DC link voltage control method.

3.1 TYPES OF CONTROLLERS:

For controlling of error in DC link voltage and higher order harmonics in source current due to nonlinear load, controllers are required. They are

1. Proportional integral controller(PI)
2. Proportional Resonant controller(PR)

3.1.1 PI CONTROLLER:

PI control is becoming one of the most studied, well known and established control techniques in most of the applications, in which the constant or slowly varying reference should be tracked. However, the application is to control alternating signals of PI regulators is not straight forward.

Application of PI controllers in stationary frame gives rise to a significant steady state error, because they only assure perfect tracking at 0 Hz. In order to achieve unit closed loop gain for alternating signals, these should be firstly transformed into dc values. This can be achieved by means of the park transform.

A conventional PI controller is defined by the transfer function

$$G_{PI}(s) = K_P + K_I/s \tag{3.2}$$

Where K_P and K_I are the proportional and integral gains, respectively.

3.1.2 PR CONTROLLER:

In case of three-phase systems, synchronous frame PI control with voltage feed forward can be used, but it usually requires multiple frame transformations, and can be difficult to implement using a low-cost fixed-point digital signal processor (DSP). Overcoming the computational burden and still achieving virtually similar frequency response characteristics as a synchronous frame PI controller, [10-11], develops the Proportional Resonant (PR) controller for reference tracking in the stationary frame. Interestingly, the same control structure can also be used for the precise control of a single-phase converter [11]. In brief, the basic functionality of the PR controller is to introduce an infinite gain at a selected resonant frequency for eliminating steady-state error at that frequency, and is therefore conceptually similar to an integrator whose infinite DC gain forces the DC steady-state error to zero. The resonant portion of the PR controller can therefore be viewed as a generalized AC integrator (GI), as proven in [12]. With the introduced flexibility of tuning the resonant frequency, attempts at using multiple PR controllers for selectively compensating low-order harmonics have also been reported in [12, 13] for three-phase active power filters, in [14] for three-phase uninterruptible power supplies (UPS) and in[15] for single-phase photovoltaic (PV) inverters. Based on similar concept, various harmonic reference generators using PR filters have also been proposed for single-phase traction power conditioners [16] and three-phase active power filters [17].

Using the PR controllers, the converter reference tracking performance can be enhanced and previously known short comings associated with conventional PI controllers can be alleviated. These short comings include steady state error in single phase systems and the need for synchronous d-q transformation in three phase systems. Based on similar control theory, PR filters can also be used for generating harmonic command reference precisely in an active power filter , especially for single phase systems, the d-q transformation theory is not applicable.

Another advantage associated with PR controllers and filters in the possibility of implementing selective harmonic compensation without requiring excessive computational resources.

The basic functionality of PR controller is to introduce infinite gain at a particular frequency for eliminating steady state error at that particular frequency. Transfer function of the PR controller is

$$G_{pr}(s) = K_p + sK_i/(s^2+\omega^2) \tag{3.3}$$

3.1.2.1DERIVATION OF SINGLE-PHASE PR TRANSFER FUNCTIONS:

For single-phase PI control, the popularly used synchronous d–q transformation cannot be applied directly, and the closest equivalence developed to date is to multiply the feedback error e(t), in turn, by sine and cosine functions usually synchronized with the grid voltage using a phase-locked-loop (PLL), as shown in Fig 3.1 [16, 23].This achieves the same effect of transforming the component at the chosen frequency to DC, leaving all other components as AC quantities.

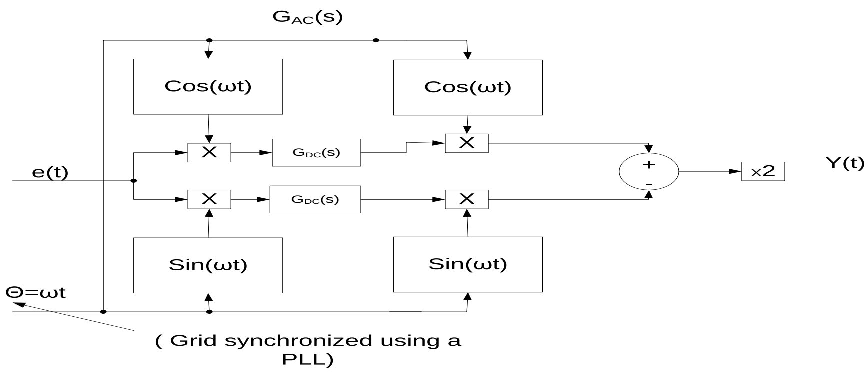

Fig 3.1 Single-phase equivalent representations of PR and synchronous PI controllers[8]

Instead of transforming the feedback error to the equivalent synchronous frame for processing, an alternative approach of transforming the controller $G_{DC}(s)$ from the synchronous to the stationary frame is also possible. This frequency-modulated process can be mathematically expressed as:

$$G_{AC}(s) = G_{DC}(s - j\omega) + G_{DC}(s + j\omega) \tag{3.4}$$

Where $G_{AC}(s)$ represents the equivalent stationary frame transfer function [16]. Therefore, for the ideal and non-ideal integrators of $G_{DC}(s) = K_i/s$ and $G_{DC}(s) = K_i/(1+(s/\omega_c))$ (Ki and $\omega_c \langle\langle \omega$ represent controller gain and cutoff frequency respectively), the derived generalized AC integrators $G_{AC}(s)$ are expressed as:

$$\begin{aligned} G_{AC}(s) &= \frac{Y(s)}{E(s)} = \frac{2k_i s}{s^2 + \omega^2} \\ G_{AC}(s) &= \frac{Y(s)}{E(s)} = \frac{2k_i(\omega_c s + \omega_c^{\,2})}{s^2 + 2\omega_c s + (\omega_c^{\,2} + \omega^2)} \\ G_{AC}(s) &\approx \frac{2k_i \omega_c s}{s^2 + 2\omega_c s + \omega^2} \end{aligned} \tag{3.5}$$

Equation (3.5), when grouped with a proportional term Kp, gives the ideal PR controller with an infinite gain at the AC frequency of ω, and no phase shift and gain at other frequencies. For Kp, it is tuned in the same way as for a PI controller, and it basically determines the dynamics of the system in terms of bandwidth, phase and gain margin. To avoid stability problems associated with an infinite gain, (3.6) can be used instead of (3.5) to give a non-ideal PR controller and, its gain is now finite, but still relatively high for enforcing small steady-state error. Another feature of (3.6) is that, unlike (3.5), its bandwidth can be widened by setting ω_c appropriately, which can be helpful for reducing sensitivity towards (for example) slight frequency variation in a typical utility grid (for (3.6), Ki can be tuned for shifting the magnitude response vertically, but This does not give rise to a significant variation in bandwidth). In passing, note that a third control structure of $G_{AC}(s) = 2k_i\omega/(s^2+\omega^2)$, can similarly be used since according to the internal model principle, it introduces a mathematical model that can generate the required sinusoidal reference along the open-loop control path, and therefore can ensure overall zero steady-state error[17].This third form is, however, not preferred since the absence of a zero at s=0 causes its response to be relatively slower[17].

Besides single frequency compensation, selective harmonic compensation can also be achieved by cascading several resonant blocks tuned to resonate at the desired low-order harmonic frequencies to be compensated for. As an example, the transfer functions of an ideal and a non-ideal harmonic compensator (HC) designed to compensate for the 3rd, 5th and 7th harmonics (as they are the most prominent harmonics in a typical current spectrum) are given as:

$$G_h(s) = \sum_{h=3,5,7} \frac{2k_{ih}s}{s^2 + (h\omega)^2}$$

$$G_h(s) = \sum_{h=3,5,7} \frac{2k_{ih}\omega_c s}{s^2 + 2\omega_c s + (h\omega)^2} \qquad (3.6)$$

Where h is the harmonic order to be compensated for and K_{ih} represents the individual resonant gain, which must be tuned relatively high (but within stability limit) for minimizing the steady-state error. An interesting feature of the HC is that it does not affect the dynamics of the fundamental PR controller, as it compensates only for frequencies that are very close to the selected resonant frequencies.

Because of this selectiveness, (3.2) with K_{ih} set to unity, implying that each resonant block now has a unity resonant peak can also be used for generating harmonic command reference in an active filter. The generic block representation is given in Fig 3.4, where the distorted load current (or voltage) is sensed and fed to the resonant filter $G_h(s)$. Obviously, the presence of unity (or 0dB) resonant peaks at only the selected filtering frequencies of 150, 250 and 350Hz for extracting the selected harmonics as command reference for the inner current loop. Also that as, ω_c gets smaller; $G_h(s)$ becomes more selective (narrower resonant peaks). However, using a smaller ω_c will make the filter more sensitive to frequency variations, lead to a slower transient response and make the filter implementation on a low-cost 16-bit DSP more difficult owing to coefficient quantization and round-off errors. In practice, ω_c values of 5–15rad/s have been found to provide a good compromise [16].

3.1.2.2 DERIVATIONS OF THREE-PHASE PR TRANSFER FUNCTIONS:

For three-phase systems, elimination of steady-state tracking error is usually performed by first transforming the feedback variable to the synchronous d–q reference frame before applying PI control. Using this approach, double computational effort must be devoted under unbalanced conditions, during which transformations to both the positive- and negative-sequence reference frames are required.

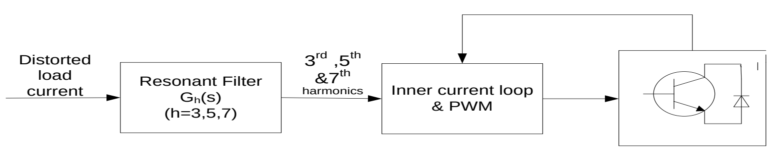

Fig 3.2 Block representation of Resonant filter for filtering 3rd, 5th and 7th harmonics [8]

Three phase proportional resonant (PR) transfer functions are used for the separation of positive and negative sequence components. In general proportional integral transfer functions are used for the separation of positive and negative sequence components but it requires transformation from stationary reference frame (αβ) to synchronous reference frame (dq), which will increase the computational effort. Therefore we use PR transfer functions in terms of stationary reference frame. Here we have to do reverse transformation from dq to αβ.

The inverse transformation can be performed by using the following 2×2matrix:

$$G_{\alpha\beta}(s) = \frac{1}{2}\begin{bmatrix} G_{dq1} + G_{dq2} & jG_{dq1} - jG_{dq2} \\ -jG_{dq1} + jG_{dq2} & G_{dq1} + G_{dq2} \end{bmatrix} \tag{3.7}$$

Where $G_{dq1} = G_{dq}(s+j\omega)$

$G_{dq2} = G_{dq}(s-j\omega)$

Given that $G_{dq}(s) = K_p + K_i/s$ and $G_{dq}(s) = K_p + K_i/(1 + (s/\omega_c))$, the equivalent controllers in the stationary frame for compensating for positive-sequence feedback error are therefore expressed as:

$$G_{\alpha\beta}^{+1}(s) = \frac{1}{2}\begin{bmatrix} K_p + \frac{2K_i s}{s^2 + \omega^2} & \frac{2K_i \omega}{s^2 + \omega^2} \\ -\frac{2K_i \omega}{s^2 + \omega^2} & K_p + \frac{2K_i s}{s^2 + \omega^2} \end{bmatrix} \tag{3.8}$$

$$G_{\alpha\beta}^{+1} \approx \frac{1}{2}\begin{bmatrix} K_p + \frac{2K_i \omega_c s}{s^2 + 2\omega_c s + \omega^2} & \frac{2K_i \omega}{s^2 + 2\omega_c s + \omega^2} \\ -\frac{2K_i \omega}{s^2 + 2\omega_c s + \omega^2} & K_p + \frac{2K_i \omega_c s}{s^2 + 2\omega_c s + \omega^2} \end{bmatrix} \tag{3.9}$$

Similarly, for compensating for negative sequence feedback error, the required transfer functions are expressed as:

$$G_{\alpha\beta}^{-1}(s) = \frac{1}{2}\begin{bmatrix} K_p + \frac{2K_i s}{s^2 + \omega^2} & -\frac{2K_i \omega}{s^2 + \omega^2} \\ \frac{2K_i \omega}{s^2 + \omega^2} & K_p + \frac{2K_i s}{s^2 + \omega^2} \end{bmatrix} \tag{3.10}$$

$$G_{\alpha\beta}^{-1}(s) \approx \frac{1}{2}\begin{bmatrix} K_p + \dfrac{2K_i\omega_c s}{s^2+2\omega_c s+\omega^2} & -\dfrac{2K_i\omega}{s^2+2\omega_c s+\omega^2} \\ \dfrac{2K_i\omega}{s^2+2\omega_c s+\omega^2} & K_p + \dfrac{2K_i\omega_c s}{s^2+2\omega_c s+\omega^2} \end{bmatrix} \quad (3.11)$$

Comparing equations (3.8) and (3.9) with (3.10) and (3.11). It is noted that the diagonal terms of $G_{\alpha\beta}^{+}(s)$ and $G_{\alpha\beta}^{-}(s)$ are identical, but there non-diagonal terms are opposite in polarity. This is inversion of polarity can be viewed as equivalent to the reversal of rotating direction between the positive and negative sequence synchronous frames.

Combining both the equations, the resulting controllers for compensating for both positive- and negative- sequence feedback errors are as expressed as:

$$G_{\alpha\beta}(s) = \frac{1}{2}\begin{bmatrix} K_p + \dfrac{2K_i s}{s^2+\omega^2} & 0 \\ 0 & K_p + \dfrac{2K_i s}{s^2+\omega^2} \end{bmatrix} \quad (3.12)$$

$$G_{\alpha\beta}(s) \approx \frac{1}{2}\begin{bmatrix} K_p + \dfrac{2K_i\omega_c s}{s^2+2\omega_c s+\omega^2} & 0 \\ 0 & K_p + \dfrac{2K_i\omega_c s}{s^2+2\omega_c s+\omega^2} \end{bmatrix} \quad (3.13)$$

Equations (3.7- 3.13) are getting from [8]

The equations (3.9) and (3.10) are used for the separation of positive and negative sequences by the PR control method. But in case of Conventional positive sequence control, we have to neglect negative Sequence control of dc link voltage.

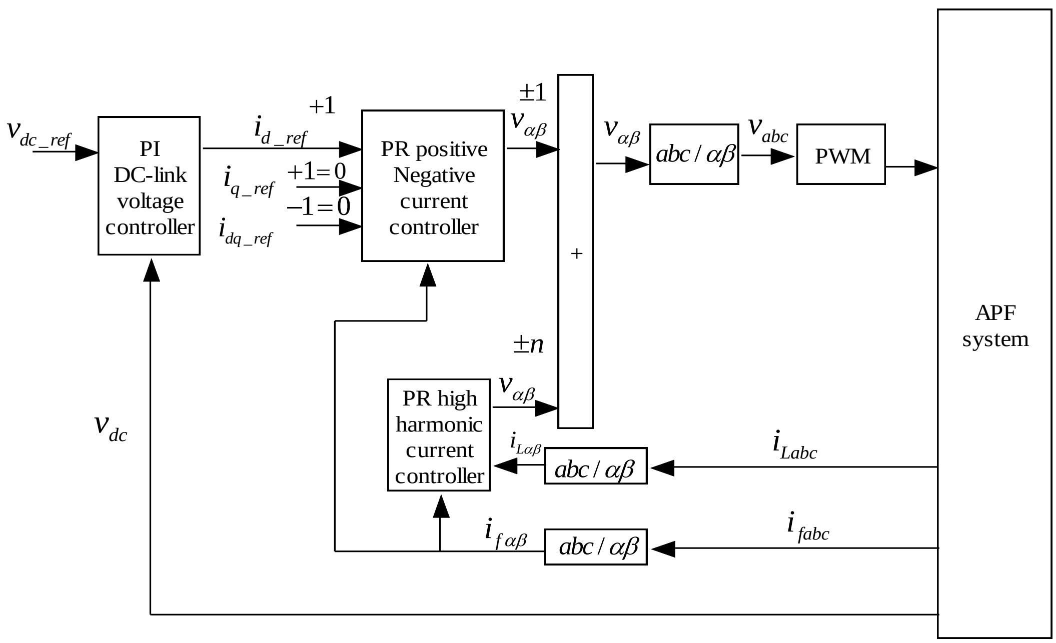

Fig 3.3 PR control block diagram for the APF system.

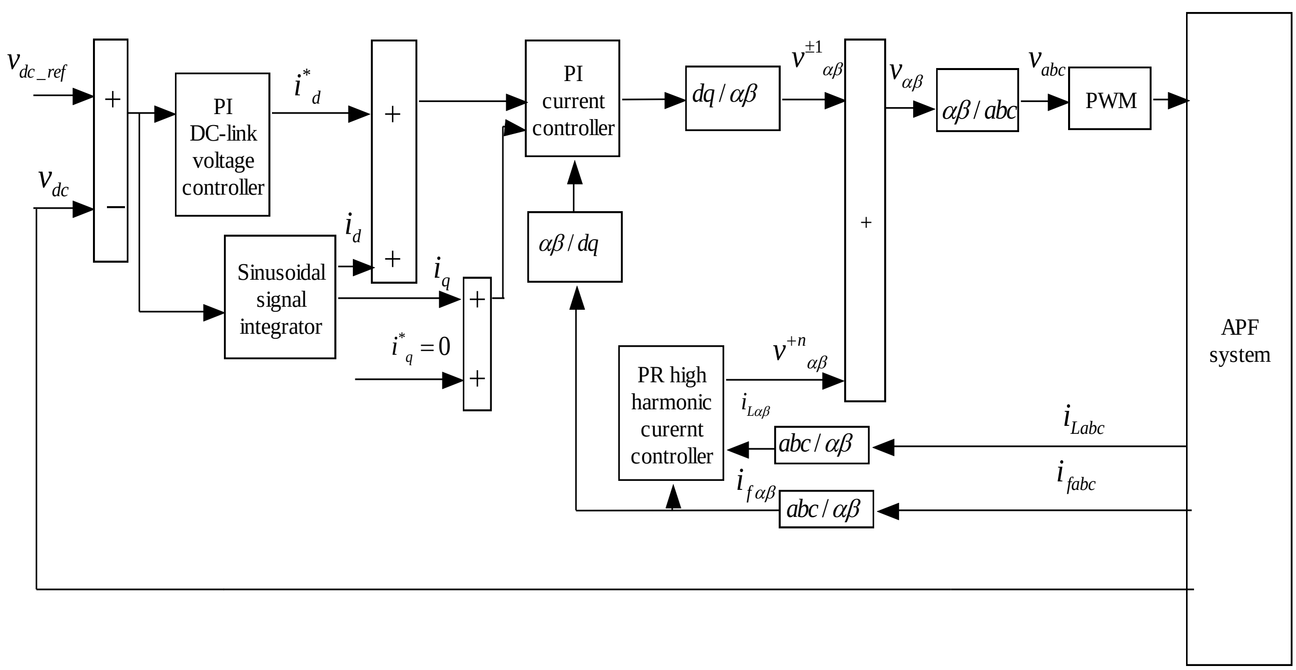

Fig 3.4 the DC link voltage control block diagram for the APF system

CHAPTER 4

RESULTS AND DISCUSSION

4. RESULTS:

4.1CONVENTIONAL POSITIVE SEQUENCE CONTROL METHOD:

In the conventional positive sequence control method positive sequence is controlled. Negative sequence is not regulated. Fig 4.1 shows the line current wave form after compensation of harmonics from the load current. Fig 4.2 shows the dc link voltage ripple due to even harmonics present in the DC link. 2^{nd} harmonic component of DC link voltage is 0.3865V. THD of line current is 2.62%. 3^{rd} harmonic component of line current is 0.1368A.

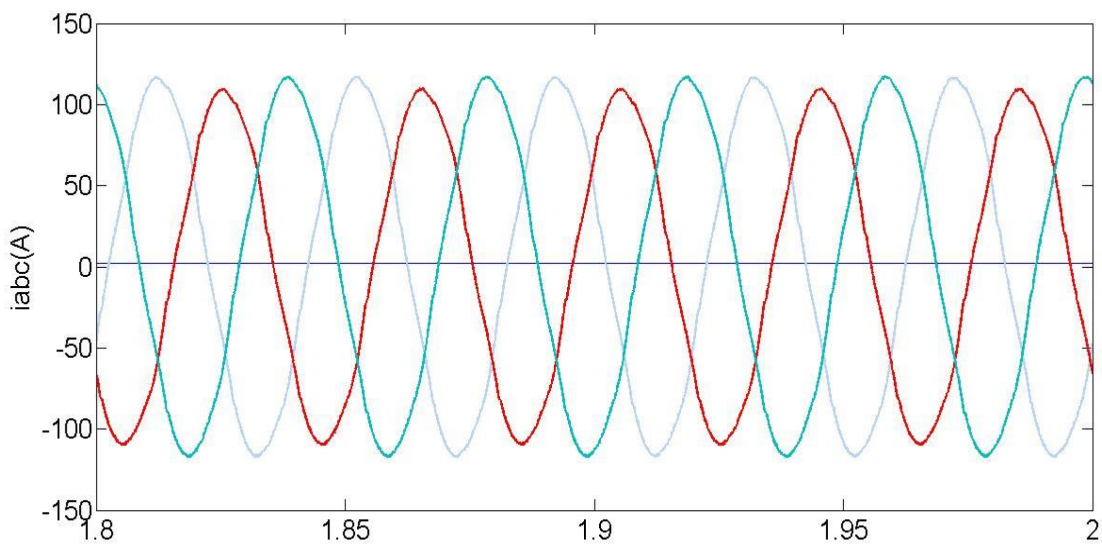

Fig 4.1 line current.

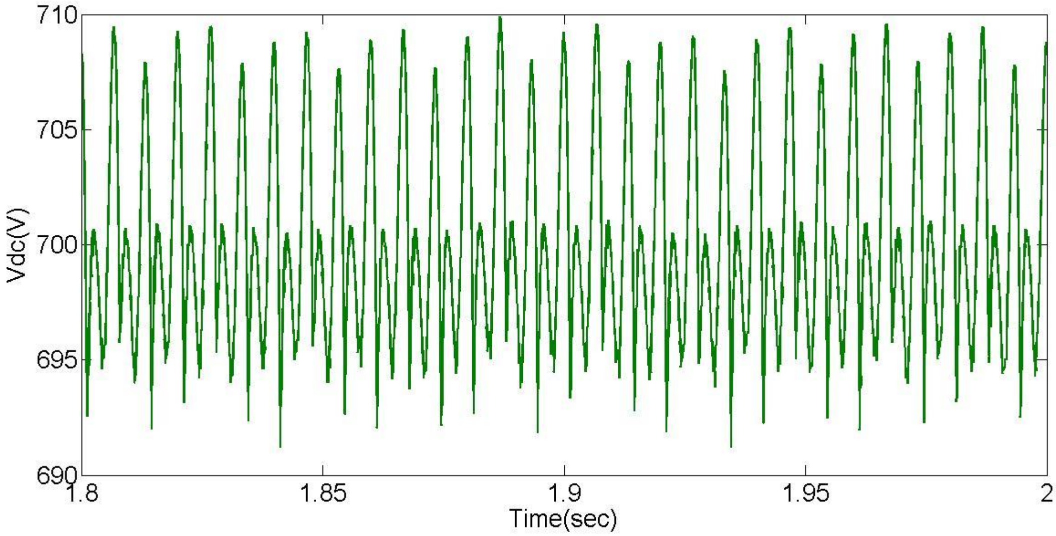

Fig 4.2 DC link voltage.

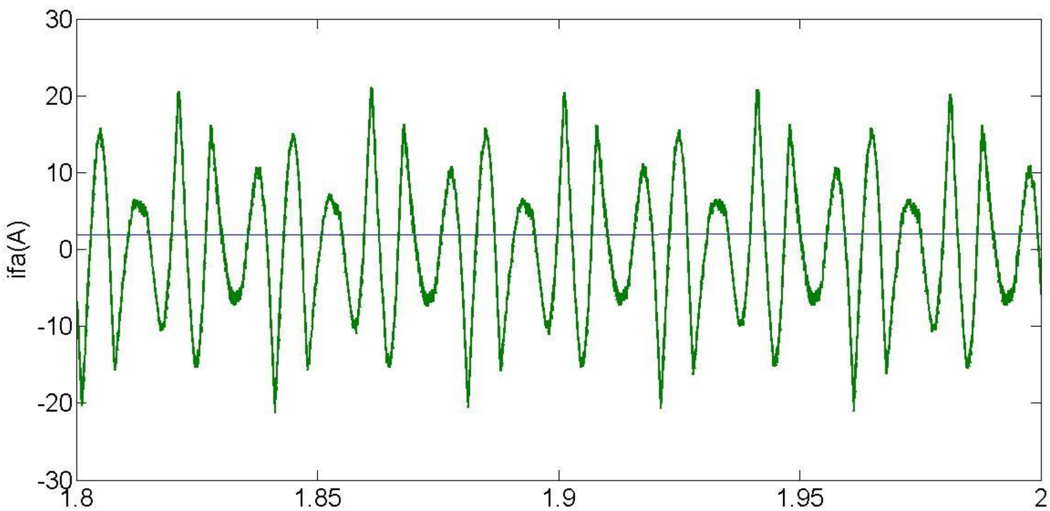

Fig 4.3 Compensating current.

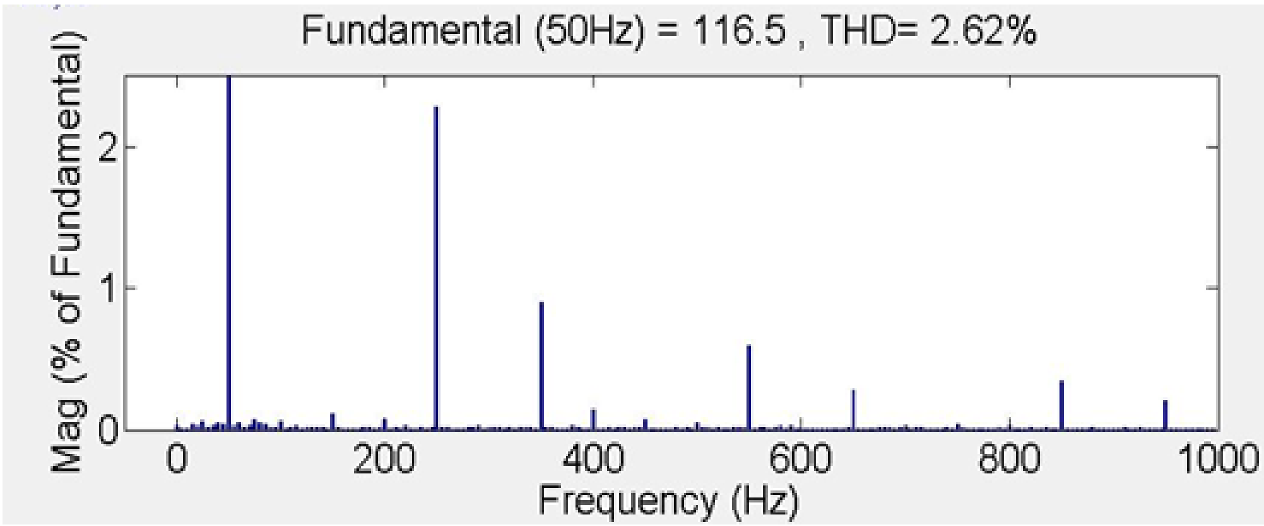

Fig 4.4 (a) THD of line current

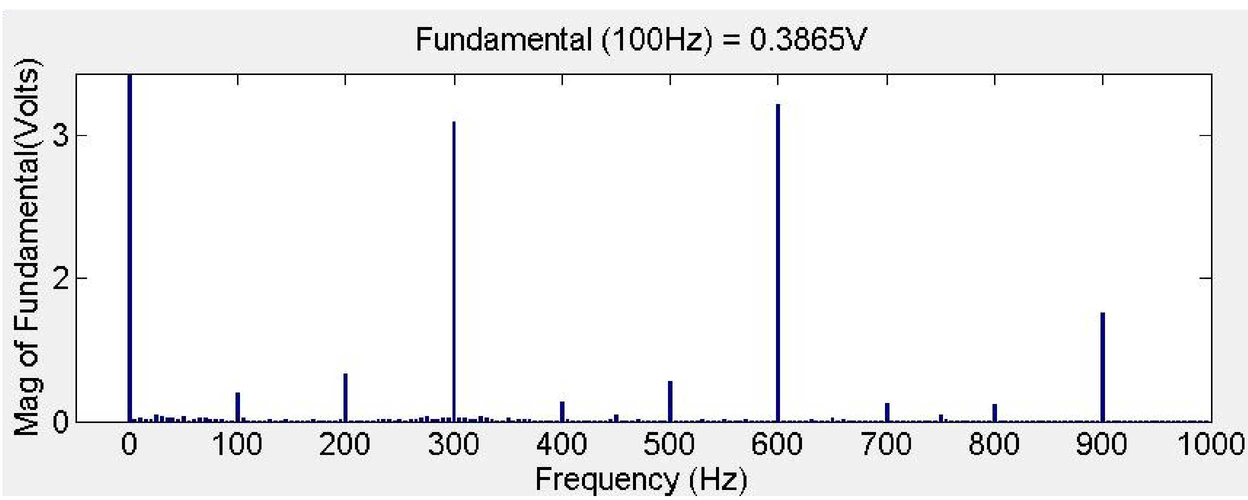

Fig 4.4 (b) Harmonic content of the DC link voltage

4.2 *DC LINK VOLTAGE CONTROL METHOD:*

This method can control positive and negative sequences. PI controllers are used for controlling of both positive and negative sequences. Fig 4.5 shows the line current wave form after compensation of harmonics from the load current. Fig 4.6 shows the dc link voltage ripple due to even harmonics present in the DC link. 2^{nd} harmonic component of DC link voltage is

0.3098V. THD of line current is 2.64%. 3rd harmonic component of line current is 0.2591A. Ripple in the DC link voltage is less comparatively conventional positive sequence method.

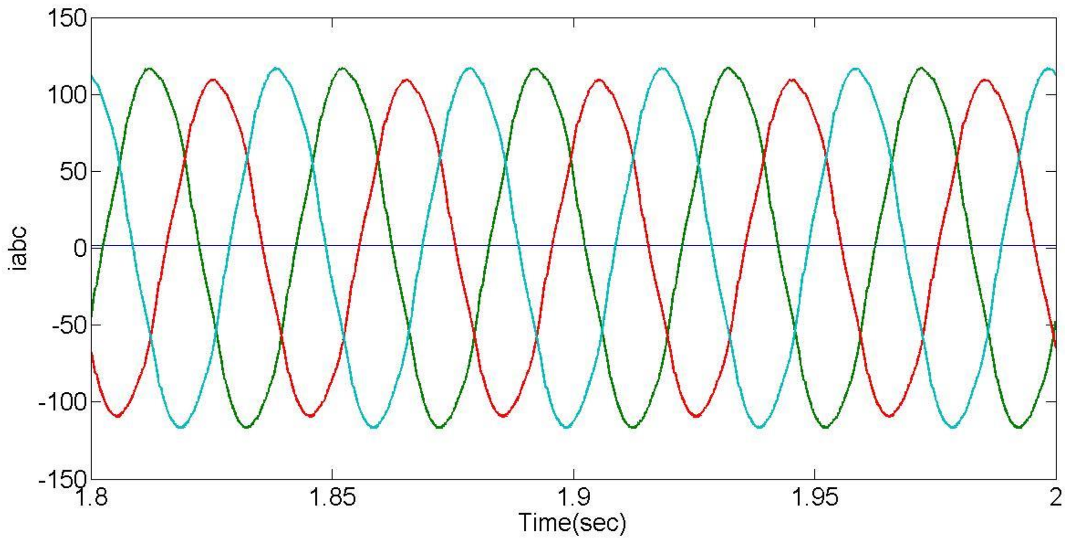

Fig 4.5 line current.

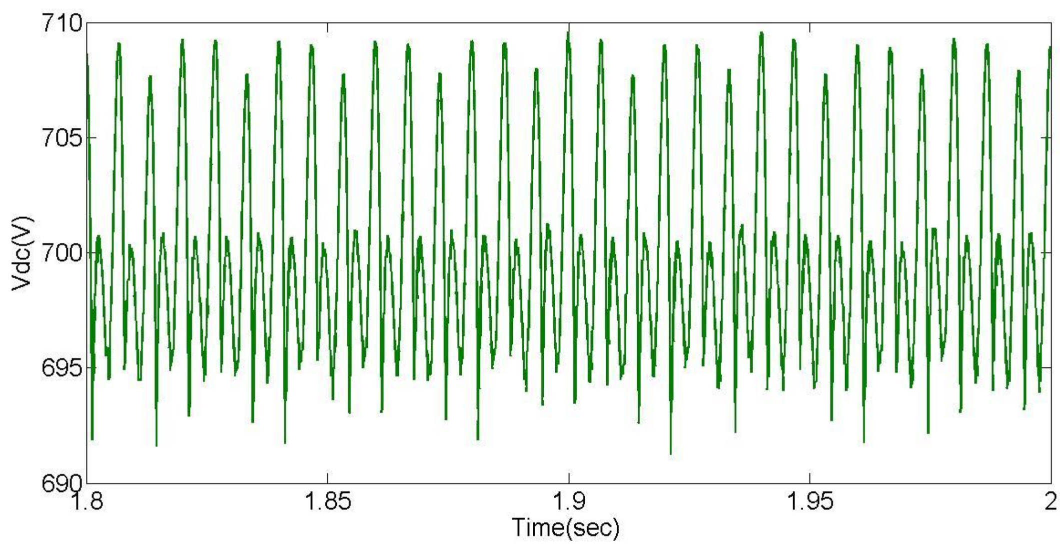

Fig 4.6 DC link voltage.

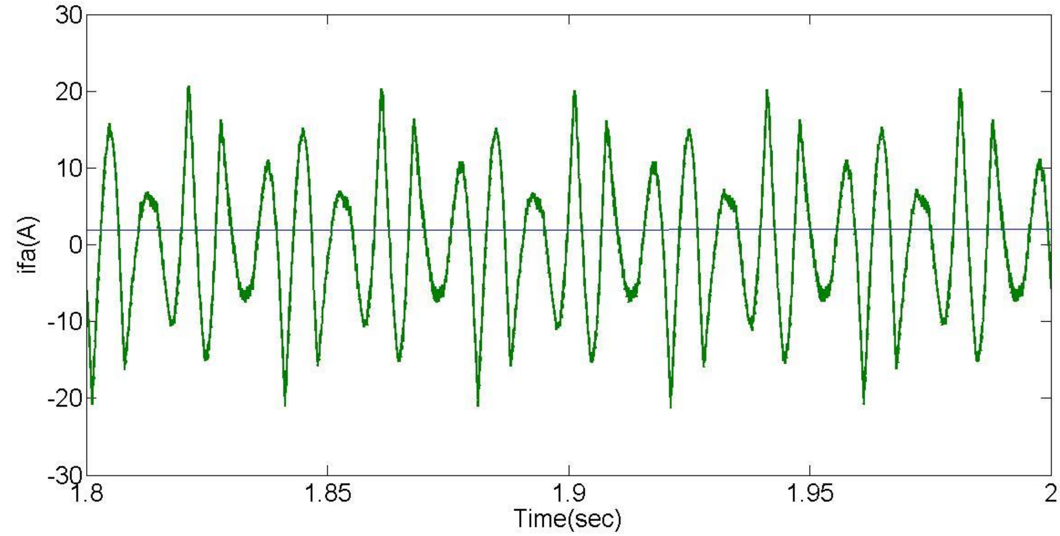

Fig 4.7 Compensating current.

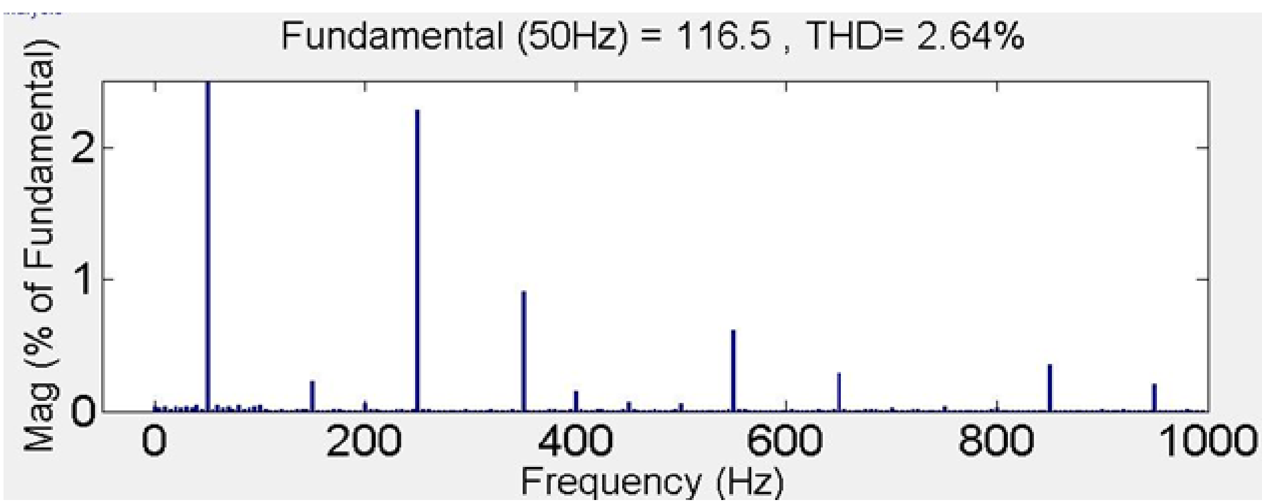

Fig 4.8 (a) THD of line current

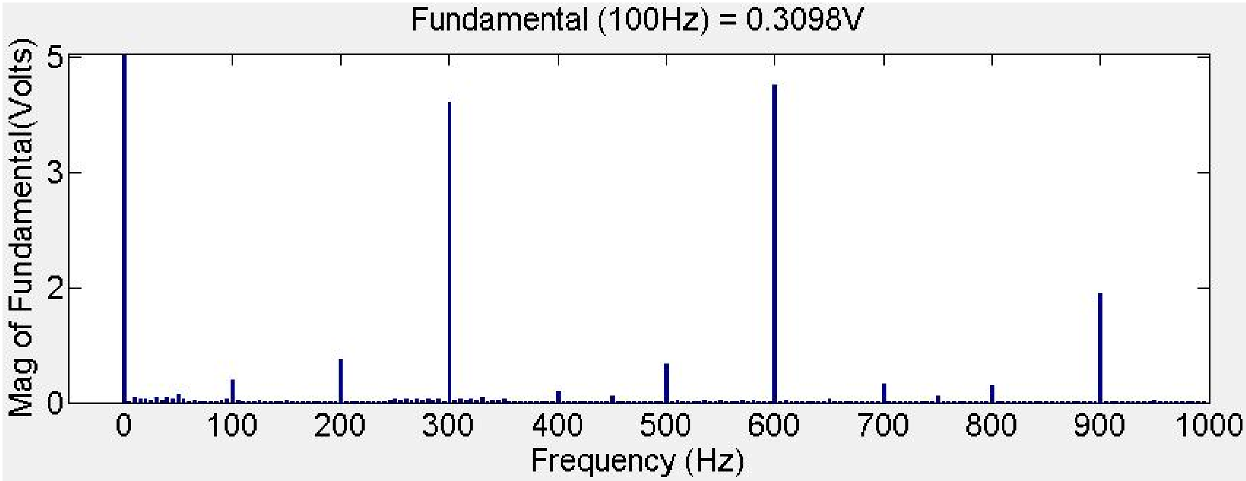

Fig 4.8 (b) Harmonic content of DC link voltage

4.3 *PROPORTIONAL RESONANT (PR) CONTROL METHOD:*

PR control method is used for the controlling of both positive and negative sequences. Fig 4.9 shows the line current wave form after compensation of harmonics from the load current. Fig 4.10 shows the dc link voltage ripple due to even harmonics present in the DC link. 2^{nd} harmonic component of DC link voltage is 0.2476V. THD of line current is 2.64%. 3^{rd} harmonic component of line current is 0.1335A. This method can reduce ripple in the DC link voltage comparatively remaining two methods.

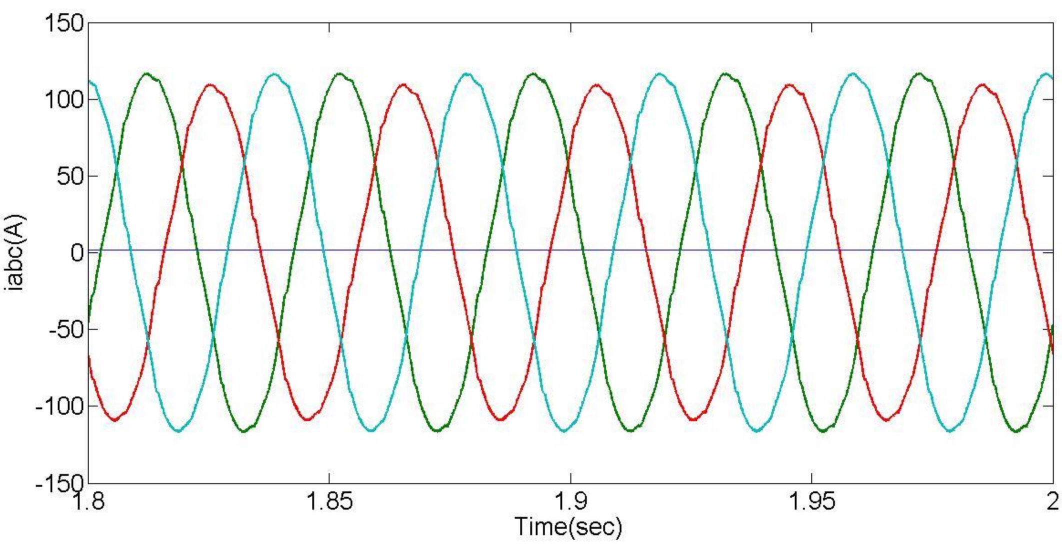

Fig 4.9 Line current.

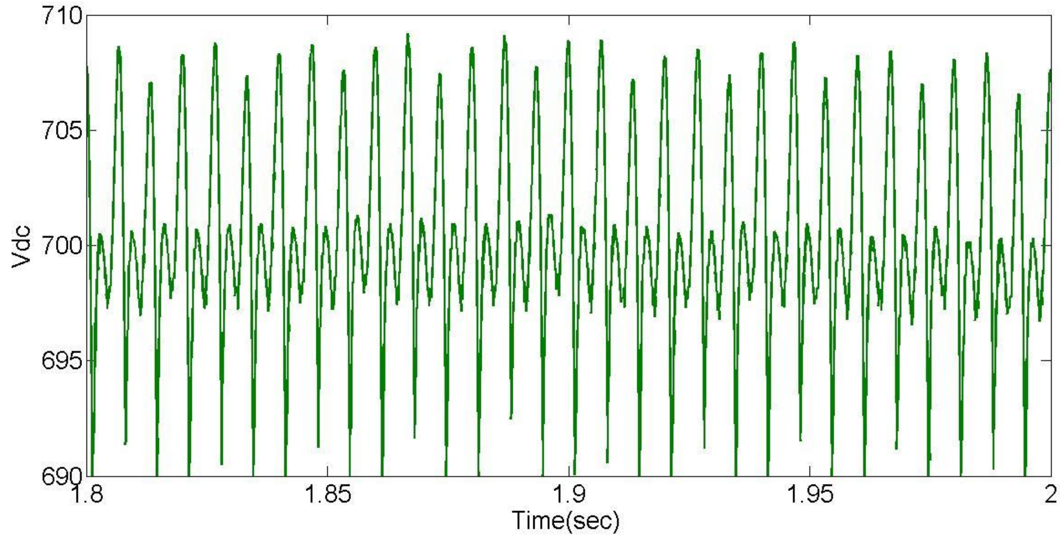

Fig 4.10 DC link voltage.

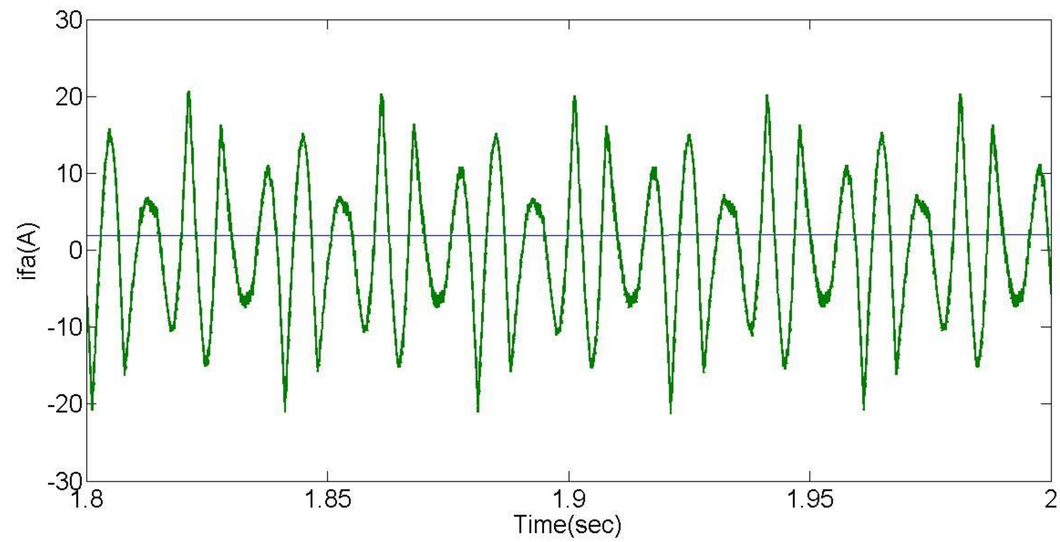

Fig 4.11 Compensating current.

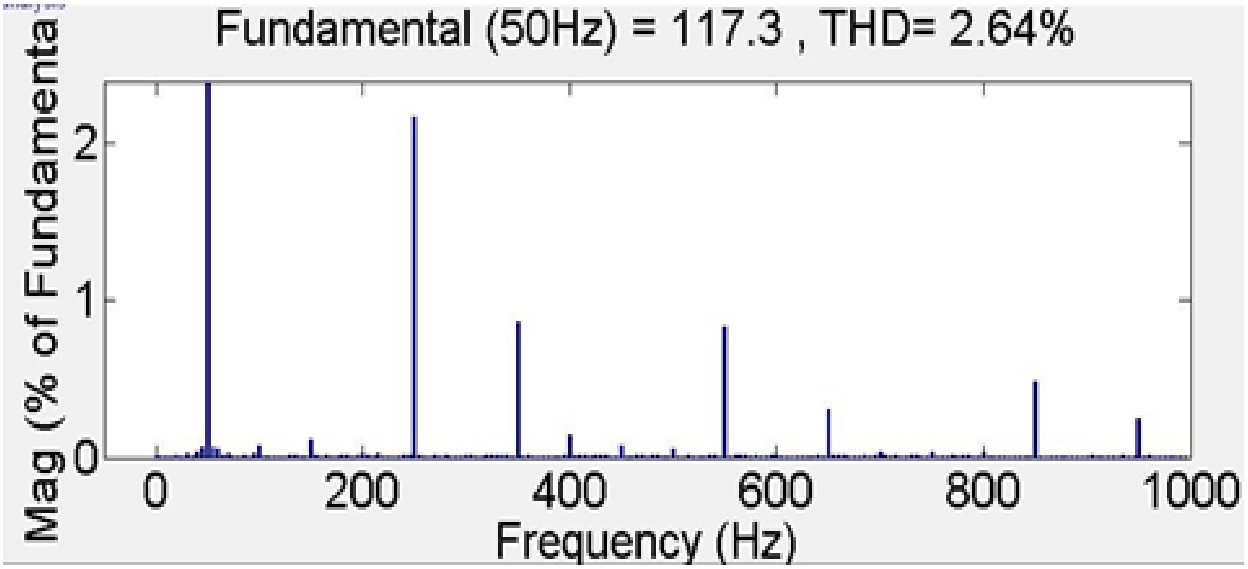

Fig 4.12 (a) THD of line current

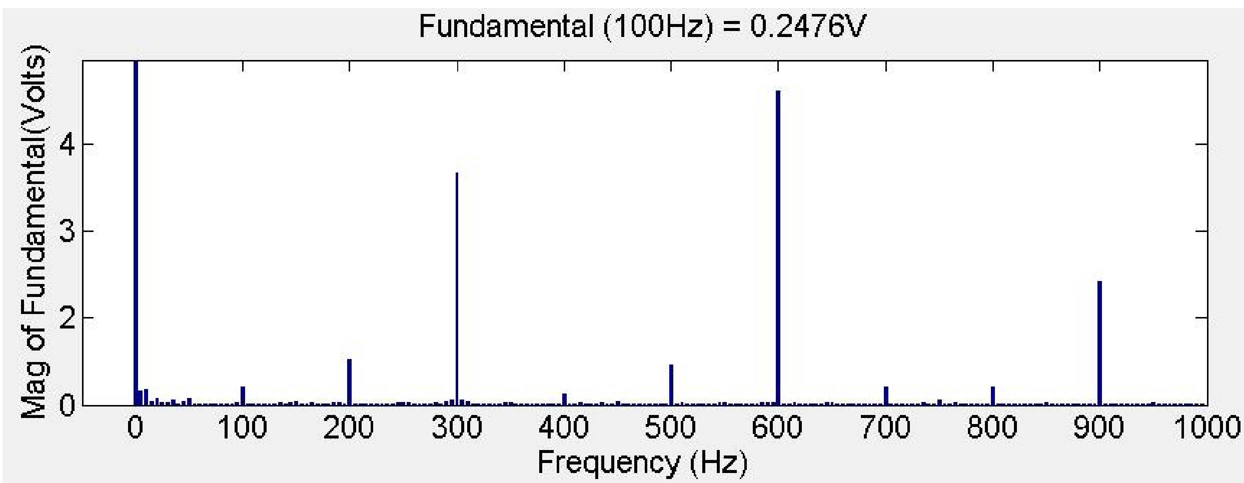

Fig 4.12 (b) Harmonic content of DC link voltage.

4.4 DISCUSSION:

Above wave forms are specified how the active power filter working under unbalanced load condition. Generally under unbalanced condition second harmonic components are injected in to the dc link voltage, shown in Fig 4.4 (b), Fig 4.8 (b) and Fig 4.12 (b). This second harmonic component will inject 3rd harmonic component in to the Ac line. Table I will show the comparison between conventional positive sequence, DC link voltage and PR control methods in the point of THD and 2nd, 4th, 6th harmonic voltages in DC link voltage.

TABLE I

Comparison between three control methods

	Conventional positive sequence control	**DC link voltage control**	**PR control**
AC current THD	2.64%	2.66%	2.64%
3rd harmonic current	0.1368A	0.2591A	0.1335A
Ripple of DC voltage (V)	1) 2nd harmonic(0.3865V) 2) 4th harmonic(0.6372V) 3) 6th harmonic(4.032V)	1) 0.3098V 2)0.5727V 3) 4.016V	1) 0.2476V 2) 0.6393V 3) 3.3V

4.5 CIRCUIT PARAMETERS:

TABLE II

Circuit parameters

Sl.no.	**Parameter name**	**value**
1	Source Voltage(phase)	220V
2	Filter inductance(la)	10mH
3	Dc link capacitance	350µF
4	Filter resistance(ra)	1mΩ
5	Dc reference Voltage	700V
6	Source inductance(Ls)	0.2856mH
7	Source Resistance(Rs)	91mΩ

CONCLUSION AND FUTURE WORK:

Comparison between conventional positive-sequence, DC link voltage control and PR control methods are discussed based on THD and 3rd harmonic component of line current and ripple of DC link voltage. The simulation results reveal that distortions caused by the 3rd order harmonics on the line current and by 2nd ripple on the DC link are nullified, combining with the conventional positive- sequence control method.

Although DC voltage control strategy cannot mitigate the 3rd order harmonic on the line current and 2nd ripple at the DC link, the PR control method is implemented more simply and cost few. This method should not use for power factor correction because it is satisfied equation (3.1). Scope of future work is to analyze current control strategies of shunt APF under unbalanced load for three phase four wire system.

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